



## An Ultra-Selective OLR – based Microstrip Diplexer with Minimal Insertion Loss for Wireless Communication System

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**Abstract:** This paper presents a high-selectivity, low-insertion-loss open-loop resonator (OLR)-based microstrip diplexer designed for wireless communication systems. We employed a microstrip transmission line with two serial capacitive gaps, incorporating rectangular half-wavelength open-loop resonators, to create a bandpass filter (BPF) with remarkable selectivity. The intended BPFs are interconnected by a T-junction combiner, which is matched to both filters and the antenna port, thereby creating the modelled diplexer. The system is implemented on a Rogers TMM4 substrate featuring a loss tangent of 0.002, a dielectric constant of 4.7, and a thickness of 1.52 mm. The suggested diplexer has dimensions of (90×70) mm<sup>2</sup>. It achieves a modest frequency spacing ratio of  $R=0.1646$  in both transmit and receive modes, with resonance frequencies of  $f_i = 2.191$  GHz and  $f_r = 2.584$  GHz, respectively. The simulated structure demonstrates favorable insertion losses of approximately 1.2 dB and 1.79 dB for the two channels at fractional bandwidths of 1.24% (2.191 GHz) and 0.636% (2.584 GHz). The simulated isolation values at 2.191 GHz and 2.584 GHz are 53.3 dB and 66.5 dB, respectively. Notably, the simulated and measured results exhibit a high degree of consistency. However, minor variations between the simulated and actual results have been noted, primarily attributable to the inherent fabrication tolerances.

**Keywords:** Microstrip diplexer, Band pass filter (BPF), Open loop resonator (OLR).

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### 1. Introduction

All wireless communication technologies are essential to our daily lives and play a significant part in many wireless system applications [1]. A key component of frequency division duplex technology is the microwave diplexer [2]. Diverse communication technologies, such as radio transmission, mobile telephony, and broadband wireless communications, all depend on diplexers.

Matthaei et al. & Wenzel conducted extensive research on diplexers in the 1960s. [3, 4]. These are very discerning parts that can combine or divide signals with various center frequencies.

Devices with three ports that divide an input signal into two channels operating at different frequencies are called diplexers [5] as shown in Fig. 1. There are various methods for creating RF diplexers, but typically three methods are utilized [6]. The first method involves using two band-pass filters with various resonant frequencies. The second

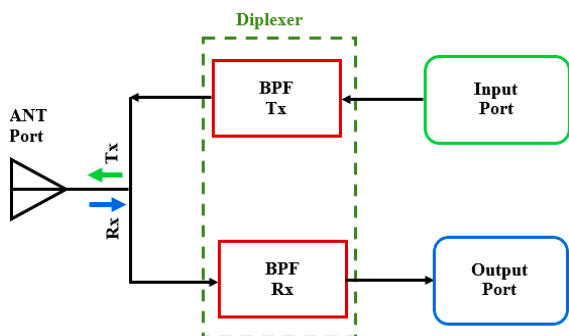


Figure. 1 Diplexer's schematic representation

method uses a band-stop filter to reject the same frequency band for one channel while using a band-pass filter for the other channel. The third method involves using two band-stop filters with different resonant frequencies.

Several types of microstrip diplexers are presented in references [7-23], but the main drawback is that they require a lot of space. Reference [7] proposes a Y-shaped slot antenna with a cavity-backed self-diplexing system using substrate integrated waveguide (SIW) to achieve high isolation. This antenna design has a Y-shaped slot on the top side of the rectangular SIW to produce two asymmetrical radiating apertures for radiation at 3.9 GHz and 4.9 GHz. While this design provides a high-gain antenna, its Tx-Rx isolation of 34 dB is not ideal, and it has a high-frequency space ratio of  $r=0.227$ . Reference [8] presents an alternative design, featuring a self-diplexing slot antenna with a bow-tie shape, incorporated within an SIW cavity. This particular design incorporates two separate feed lines to induce resonance within the SIW cavity at both the transmitting and receiving frequencies, yielding a high-gain antenna exhibiting a unidirectional radiation pattern. Nonetheless, it is important to note that this design has a relatively large high-frequency space ratio ( $r=0.22$ ) and exhibits limited transmit-receive isolation, measuring at 22 dB.

In reference [9], a microstrip diplexer is employed to merge the global mobile system (GSM) and wireless local area network (WLAN) frequency bands. The negative effects of the microstrip diplexer are reduced by this design. It has some shortcomings, such as significant channel losses and a 30 dB isolation limit. A straightforward method for designing a microstrip diplexer is suggested in [10]. It is built using two small Square OLR band pass filters together. These filters were created for 2.45 GHz Radio Frequency Identification applications. Chebychev's approximation is used. Diplexer isolation is 40 dB.

A dual-mode resonator-based substrate integrated waveguide (SIW) diplexer is suggested in [11]. This

diplexer is used to enhance RF front end performance. This diplexer provides appropriate isolation of 49 dB and 53 dB, respectively, for the broadcast and receive channels. Yet, some of its shortcomings are its size.

In reference [12], there is a proposal for the creation of a microstrip diplexer using the design of squared open-loop resonators (SOLRs). This diplexer exhibits a narrow frequency spacing with  $r=0.114$ , and it features two resonance frequencies: 1.81 GHz for the transmit channel and 2.03 GHz for the receive channel. Despite its advantages, the diplexer's dimensions are seen as a drawback., respectively. Despite its benefits, the diplexer's size is considered a limitation.

In reference [13], a dual-band bandpass filter was created by merging two independent channel filters to establish a microstrip diplexer. This approach eliminates the requirement for external junctions during the diplexer's construction, which diverges from conventional techniques that demand separate connections or junctions for distributing energy. Through simulations and experimental observations, the diplexer demonstrated an impressive 50 dB isolation between its transmit and receive bands. Nevertheless, there were discernible insertion losses of 2.88 dB in the transmitting band and 2.95 dB in the receiving band.

In another work referenced as [14], a new shaped microstrip diplexer with improved isolation and selectivity was introduced. This design employs two compact bandpass filters comprised of open/shorted lines and an open stub, tailored for applications related to LTE. For the frequency response performance to be improved, many types of resonators are used [15-20]. Recently, other resonator shapes such as U-shaped [15], Triangular loop resonators [16], T-shaped, [17], have been introduced. Pi-shaped [18], stepped-impedance [19], and Patch resonators [20,21]. Patch resonators are utilized in [22] to produce a filtering response. In order to create a microstrip diplexer, two bandpass filters (BPFs) made of spiral cells and coupled lines are integrated in [23]. It has some drawbacks, including unwanted harmonics and large losses at both channels. The configuration of the showcased microstrip diplexer in reference [24] employs a setup involving three interconnected line structures to enhance the attenuation of harmonics. Nonetheless, the challenge of notable signal losses in both channels persists. On the other hand, the proposed diplexer outlined in reference [25] offers the potential to efficiently mitigate harmonic distortions and also lower the simulated insertion losses for both channels. However, both channels exhibit substantial common port return losses in this arrangement.

Table 1. Detailed comparison of latest technology of microstrip diplexer for wireless communication systems with the existing literature

#	Reference	Size $\lambda_g^2$	Frequency (GHz)	FBW (%)	Insertion Loss (dB)	Isolation (dB)	R
1	[9]	0.35	1.8/2.4	6/5.8	NA	30	0.286
2	[10]	1.2	2.2/2.6	4.55/5	1.6/1	40	0.167
3	[11]	4.6	8.04/9.07	4.23/4.19	2.35/2.33	49/53	0.12
4	[12]	1.5	1.81/2.03	2.25/3	1.98/1.9	58/46	0.114
5	[14]	0.167	1.8/2.1	5.5/6.2	2/1.8	40	0.154
6	[22]	0.89	1.85/2.5	NA	4.2/3.3	31	NA
7	[23]	0.98	1.41/2.02	NA	4.4/4.6	29	NA
8	[25]	0.3	1.5/2.1	NA	1.5/1.8	45	NA
9	[26]	3.5	9.5/10.6	6	2.6/2.3	45	NA
This work		0.435	2.191/2.584	1.24/0.63 6	1.2/1.79	53.3/66.5	0.1646

In contrast to the simulated insertion losses at both channels in [26], which are relatively low, the introduced diplexer in [27] has substantial losses at both channels. The suggested structure, however, was unable to achieve modest return losses at either its up-per or lower channels.

As endeavors are underway to develop a compact microstrip bandpass diplexer with high-performance characteristics, the design aspects, such as low losses and increased fractional band-width, make it difficult to sacrifice the compact size.

This work introduces a high isolation and low insertion loss effective design scheme for microstrip diplexers. The two-compact size BPFs that make up the proposed diplexer are constructed from linked OLRs. A T-junction that serves as a combining circuit connects the two BPFs to the antenna and provides good isolation between the up-link and down-link BPFs. Initially, a high selectivity BPF is created for 2.191 GHz operation. The suggested microstrip diplexer is then made up of the other BPF, which is created to operate at 2.584GHz, and the first BPF.

The suggested diplexer exhibits excellent size-related characteristics and performance., resonance frequency (GHz) at transmit channel and receive channel, (FBW (GHz)), insertion loss (dB), isolation (dB), and frequency space ratio (R), compared to #1 up to #9. Table 1 presents a comparative analysis of various diplexer designs, showcasing a competitive

version among the options available.

The critical contributions of this paper are listed below:

- 1) A compact diplexer with high selectivity, fractional bandwidths of 1.24% and 0.636%, low insertion losses of 1.2 dB and 1.79 dB, and a low frequency space ratio of 0.1646 was proposed.
- 2) A miniaturized two-bandpass filters (BPFs) were obtained, with the first one for the transmit channel and the second for the receive channel. The total physical dimension of the filters was 25 mm x 25 mm x 1.52 mm, with a resonance frequency of 2.191 GHz and 2.584 GHz for the transmit and receive channels, respectively.
- 3) A diplexer was obtained by assembling the two BPFs in an orthogonal manner through a T-junction to achieve the highest isolation of 53.3 dB and 66.5 dB for the transmit and receive channels, respectively.
- 4) An LC equivalent circuit and numerical formula were proposed for the BPF and diplexer to comprehend their behavior.

## 2. Proposed OLR-based microstrip diplexer

Generally, two band pass filters are constructed as the initial step in the microstrip diplexer design process. In this section, an alternative design is introduced for a highly efficient microstrip diplexer

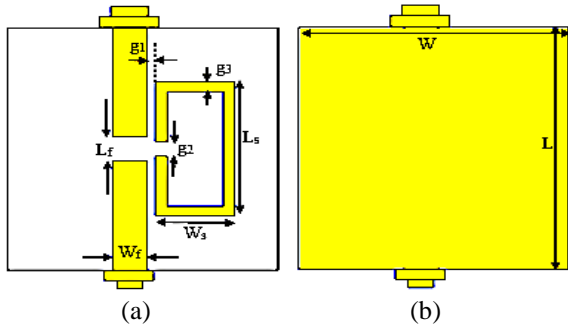


Figure. 2 The geometrical construction of the OLR- BPF at  $f_t = 2.191\text{GHz}$  for both: (a) front and (b) back views

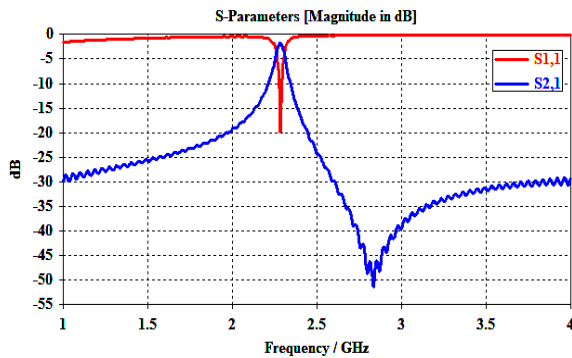


Figure. 3 The receive BPF's simulated S-parameters

that operates at  $f_i=2.191$  GHz for transmit mode and  $f_r=2.584$  GHz for receive mode, is shown using the computer simulation technology (CST) software.

The Rogers TMM4 substrate, as mentioned earlier, hosts the diplexer design on both of its surfaces. In order to achieve the desired transmit and receive frequencies,  $f_i= 2.191$  GHz and  $f_r= 2.584$  GHz, respectively, two selective BPFs are joined together to form the proposed diplexer. The T-junction is used to join the two BPFs together. As a result, the suggested diplexer's design process can be completed in the following two steps: (i) The transmit and receive BPF designs. (ii) The planned microstrip diplexer's meeting.

### 2.1 Creation of the transmit and receive bandpass filters (BPFs)

This section describes the designs of the transmit and receive bandpass filters (BPFs). Fig. 2, illustrates the proposed topology for the microstrip BPF, which includes two input/output feed lines and a rectangular open-loop resonator. The interaction between the two transmission lines and the resonator is achieved through gap  $g_1$ , with gap  $L_f$  serving as the separator between the two feed lines. The incorporation of a folded microstrip resonator, along with strategic positioning of the two feed lines, contributes to the enhanced compactness of this design.

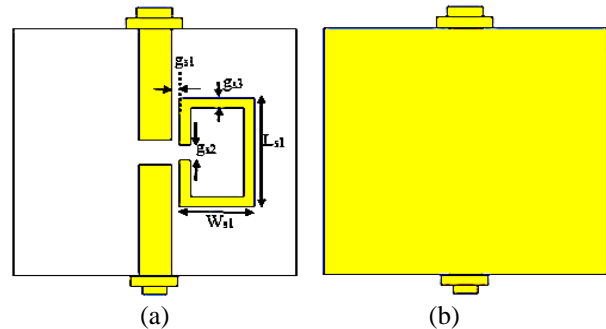


Figure. 4 The geometrical construction of the OLR-based BPF at  $f_t = 2.584$  GHz for both: (a) front and (b) back perspectives

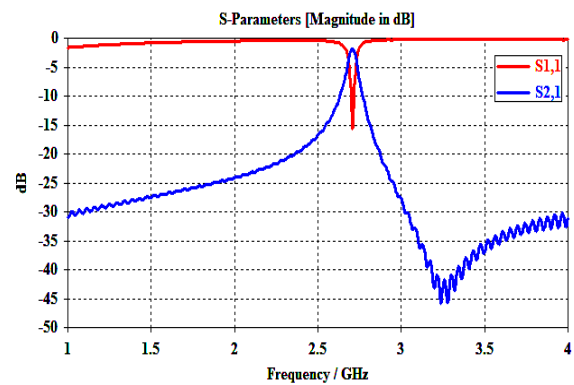


Figure. 5 The transmit BPF's simulated S-parameters

Table 2. The suggested transmit and receive BPF's dimensions

Parameter	Value (mm)	Parameter	Value (mm)
$W$	25	$L_{s1}$	11.2
$L$	25	$g_1$	0.75
$W_f$	3.06	$g_2$	1.5
$L_f$	2.5	$g_3$	1
$W_s$	7	$g_{s1}$	0.75
$L_s$	14	$g_{s2}$	1.75
$W_{s1}$	7	$g_{s3}$	1

CST-MWS-2019 was used to investigate the operation of the BPF and simulate the shown OLR-BPF, which are printed on a 1.52mm Rogers TMM4 substrate. At the transmit frequency of  $f_i= 2.191\text{GHz}$ , OLRs are used, as shown in Figure 2 each with a total length of roughly  $\lambda_g/2$ . Table 2 contains a list of the filter's dimensions. The width and length of the trace line, coupled with the separation gap  $g_1$ , determine each resonator's internal capacitance, which in turn influences the filter's selectivity and insertion loss.

The CST Microwave Studio was used to simulate the scattering parameters for the proposed transmit

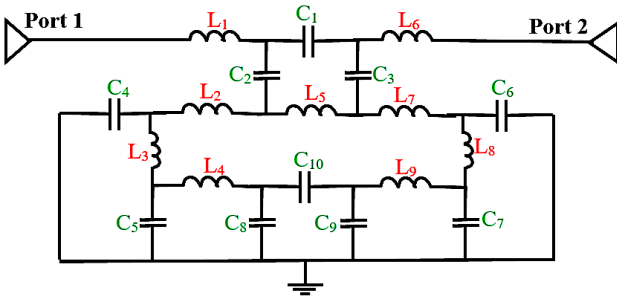


Figure. 6 The proposed BPF's LC equivalent circuit

bandpass filter (BPF), as shown in Fig. 3. This BPF operates at a center frequency of 2.191 GHz and boasts a 3 dB bandwidth spanning 47.1 MHz, with FBR of 1.24%, a return loss of 20 dB, and an insertion loss of 1.4 dB. These parameters align with a range of specified criteria. The suggested receive BPF, which has dimensions indicated in Table 2, is depicted in Fig. 4, and follows the same conceptual framework. Fig. 5, displays the receive BPF's simulated scattering parameters. The curves' analysis reveals the BPF's center frequency to be 2.584 GHz, 3dB bandwidth to be 20.7 MHz, fractional bandwidth to be 0.636%, return loss to be 15.6 dB, and insertion loss to be 1.3 dB.

In the second approach within this study, we focus on analyzing the proposed topology by transforming the microstrip structure into its corresponding LC model, as depicted in Fig. 6. This transformation is carried out with the aim of obtaining a deeper insight into the performance of the suggested BPF.

The proposed topology is symmetric, which enables the use of half circuit LC analysis to convert and transforming the microstrip transmission lines into the LC model, as depicted in Fig. 6. In this LC model, the microstrip transmission lines are assumed to be lossless, while the corresponding LC model is an approximation. The inductances of the central transmission lines are represented by  $L_1, L_2, L_3, L_4, L_5, L_6, L_7, L_8,$  and  $L_9$  in this LC model. Within the model,  $C_1$  is utilized to denote the capacitive impact of the space between the two feed-lines. Meanwhile,  $C_2$  and  $C_3$  are assigned to signify the capacitive effects of the gaps between the feed line and the resonator. Furthermore,  $C_8$  and  $C_9$  are attributed to encapsulate the capacitive consequences of the open stubs concerning the ground. Lastly, the capacitive ramifications of the bends are symbolized by  $C_3, C_4, C_5,$  and  $C_6$ . Every component of the microstrip line is then transformed to an equivalent LC lumps after being divided into smaller sections to assist the conversion process. The electrical parameters of each component are derived using the calculation method detailed in the section below, as referenced in [28].

Lastly, certain tweaking approaches are used to improve the match between the simulation results of the two circuits. The Eqs. (1) and (2) are applied in order to determine the values of each element in the LC configuration.

$$l = CZ_c v_p \quad (1)$$

$$l = \frac{Lv_p}{Z_c} \quad (2)$$

If we maintain a consistent line width, we employ the variables  $C$  and  $L$  to signify the length, capacitance, and inductance of a microstrip line. The characteristic impedance is indicated as  $Z_c$ , and the phase velocity is denoted as  $v_p$ .

$Z_c$  and  $v_p$  must first be determined in order to calculate the  $C$  and  $L$  values. When the effective dielectric constant  $\epsilon_{re}$  has been established, the phase velocity is derived by:

$$v_p = \frac{c}{\epsilon_{re}} \quad (3)$$

The symbol  $c$  presents the speed of light in a vacuum. The equations below can be utilized to determine the characteristic impedance  $Z_c$  and effective dielectric constant  $\epsilon_{re}$ :

For  $W_f/h \leq 1$

$$\epsilon_{re} = \frac{\epsilon_r+1}{2} + \frac{\epsilon_r-1}{2} \left\{ \left( 1 + 12 \frac{h}{W_f} \right)^{-0.5} \right\} + 0.04 \left( 1 - \frac{W_f}{h} \right)^2 \quad (4)$$

$$Z_c = \frac{\eta}{2\pi\sqrt{\epsilon_{re}}} \ln \left[ 8 \frac{h}{W_f} + 0.25 \frac{W_f}{h} \right] \quad (5)$$

For  $W_f/h \geq 1$

$$\epsilon_{re} = \frac{\epsilon_r+1}{2} + \frac{\epsilon_r-1}{2} \left( 1 + 12 \frac{h}{W_f} \right)^{-0.5} \quad (6)$$

$$Z_c = \frac{\eta}{\sqrt{\epsilon_{re}}} \left[ \frac{h}{W_f} + 1.393 + 0.677 \ln \left\{ \frac{W_f}{h} + 1.444 \right\} \right]^{-1} \quad (7)$$

Where  $W_f$  is the microstrip line's width,  $h$  and  $\epsilon_r$  are the substrate's thickness and permittivity,  $\epsilon_{re}$  is its effective permittivity, and the value  $\eta = 120\pi\Omega$  symbolizes the characteristic impedance in a vacuum. Each right-angle bend within the microstrip open-loop resonator can be modelled using an analogous T-network configuration, as illustrated in reference [29]. The formulas (8) and (9) provide a roughly form



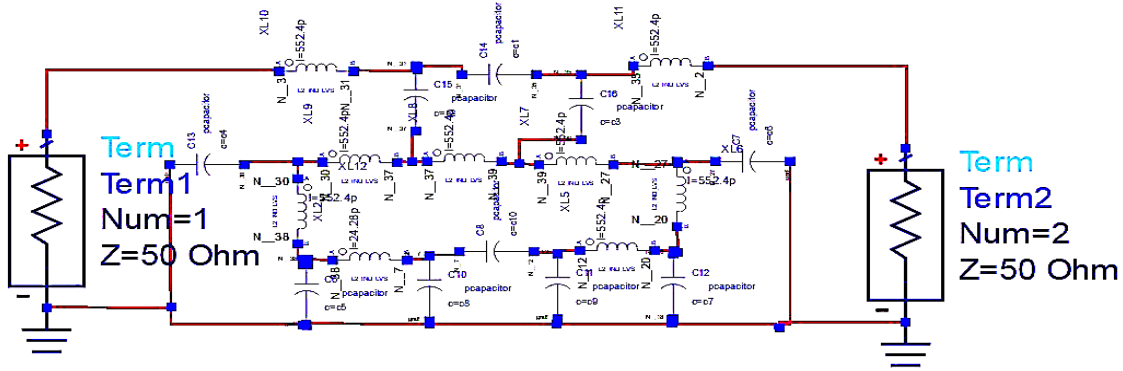


Figure. 7 The schematic circuit model of the proposed BPF in ADS simulation program

for calculating the capacitance in this work.

$$\frac{C}{W_f} (PF/m) = \frac{(14\epsilon_r + 12.5)W_f/h - (1.83\epsilon_r - 2.25)}{\sqrt{W_f/h}} + \frac{0.02\epsilon_r}{W_f/h} \text{ for } W_f/h < 1 \quad (8)$$

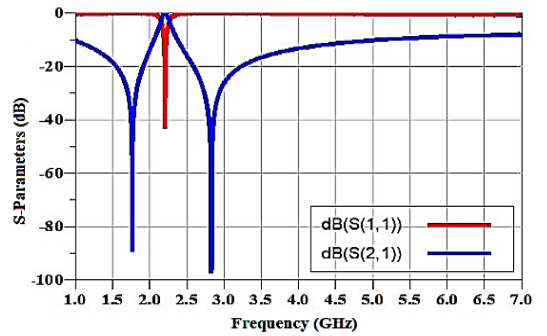
$$\frac{C}{W_f} (PF/m) = (9.5\epsilon_r + 1.25)W_f/h + 5.2\epsilon_r + 7 \text{ for } W_f/h > 1 \quad (9)$$

The effective dielectric constant and the characteristic impedance are the two primary parameters that should be determined, therefore for a line width  $W_f = 3.06mm$ , we obtain ( $\epsilon_{re} = 3.55$ ,  $Z_c = 49.9$ ). The modified circuit is then simulated in a schematic environment, following the use of the calculating procedure.

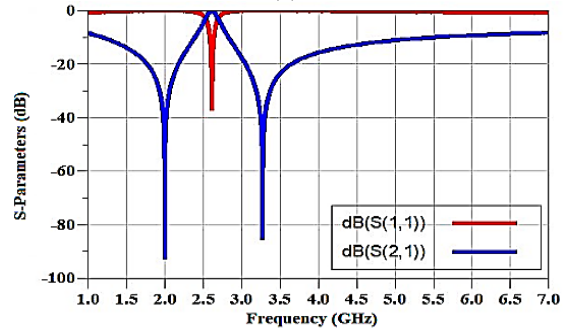
Several simulations were conducted, adjusting the values of each variable in the LC lumps to achieve a closer alignment between the electromagnetic (EM) simulation outcomes and the converted circuit results. The LC-based equivalent circuit of the suggested bandpass filter (BPF) underwent simulation using the ADS electromagnetic simulator.

The proposed BPF's schematic circuit model is shown in Fig. 7, using the ADS simulation tool. The simulated scattering parameters of the LC model using ADS are shown in Fig. 8. It is noticed that the behavior of the BPFs is ideal and allows to pass frequency 2.191GHz for transmit channel as shown in Fig. 8(a), at values of the LC components as follow:

$L_1 = 0.01$ ,  $L_2 = 0.01$ ,  $L_3 = 0.01$ ,  $L_4 = 0.01$ ,  $L_5 = 0.01$ ,  $L_6 = 0.01$ ,  $L_7 = 0.009$ ,  $L_8 = 0.01$ ,  $L_9 = 0.01$ . (all in nH).  $C_1 = 4.2$ ,  $C_2 = 1e-6$ ,  $C_3 = 0.18$ ,  $C_4 = 1e-6$ ,  $C_5 = 0.12$ ,  $C_6 = 0.12$ ,  $C_7 = 0.15$ ,  $C_8 = 7.4$ ,  $C_9 = 6.0$ ,  $C_{10} = 0.12$ . (all in pF), and 2.584 GHz for receiving channel as shown in Figure 8b at values of the LC components as follow:



(a)



(b)

Figure. 8 Scattering parameter of proposed BPF equivalent circuit in ADS simulation program: (a) At 2.191GHz, and (b) At 2.584GHz

$L_1 = 0.01$ ,  $L_2 = 0.01$ ,  $L_3 = 0.01$ ,  $L_4 = 0.01$ ,  $L_5 = 0.01$ ,  $L_6 = 0.01$ ,  $L_7 = 0.009$ ,  $L_8 = 0.01$ ,  $L_9 = 0.01$ . (all in nH).  $C_1 = 4.8$ ,  $C_2 = 1e-6$ ,  $C_3 = 0.17$ ,  $C_4 = 1e-6$ ,  $C_5 = 1e-6$ ,  $C_6 = 8.6$ ,  $C_7 = 1.4$ ,  $C_8 = 7.4$ ,  $C_9 = 6$ ,  $C_{10} = 0.12$  (all in pF).

Variation of some of the proposed structure's geometric characteristics is one of the key researches to comprehend the behaviour of the suggested band pass filter. This represents an intriguing method to enhance the filter's performance. It was discovered that a crucial design element was the distance  $L_f$  between the two feed lines. Many simulations were completed in this context to explore the frequency responses of the filter at various gap  $L_f$  values.

The results from the simulation of the  $S_{21}$

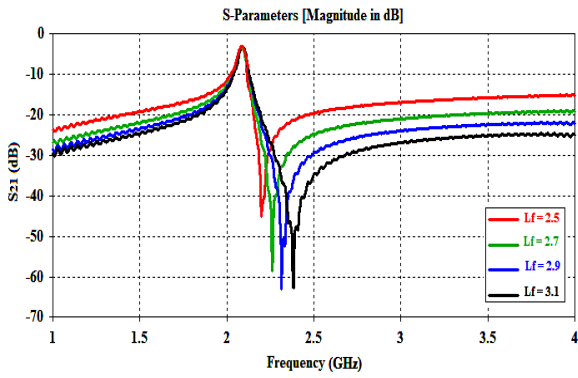


Figure. 9 The simulated S21 parameter at various gap  $L_f$  value

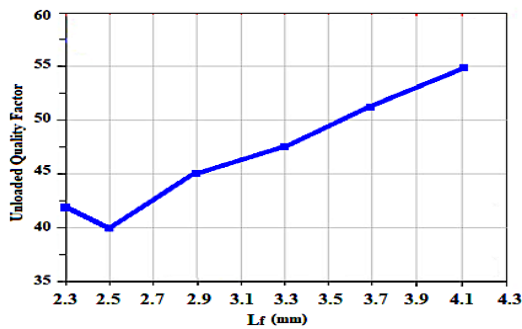


Figure. 10 The progression of the quality factor without a load in the simulation as it varies with different values of gap  $L_f$

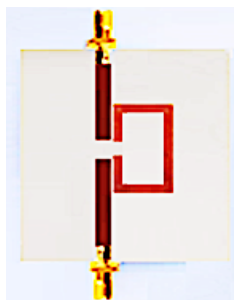


Figure. 11 Manufactured prototype of the suggested BPF

parameter for various space sizes are presented in Fig. 9. The figure reveals two notable outcomes. The first consequence pertains to the bandwidth, where an increase in gap width causes a slight increase in insertion losses and a reduction in bandwidth due to a smaller area between the feed lines and the resonator's periphery. Meanwhile, changing the gap width can impact the position and weakening of the transmission zero, as shown in the figure. The position and weakening of the transmission zero increase as the gap width widens, indicating enhanced selectivity. The existence of the transmission zero results from the cross-coupling between the two resonator arms and the two input/output lines. Hence, modifying the space size can account for the observed behaviour.

Modifying the gap size also impacts the return

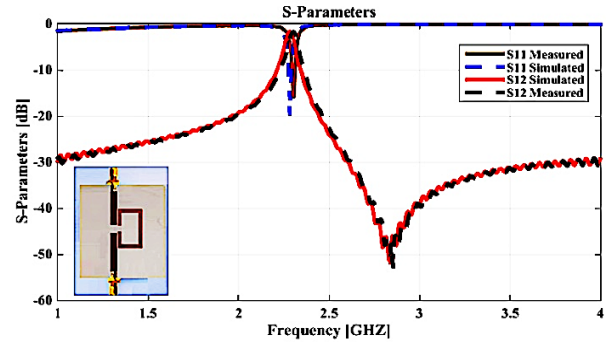


Figure. 12 Evaluating the scattering parameters of the designed bandpass filter (BPF) through a comparison of measured and simulated results

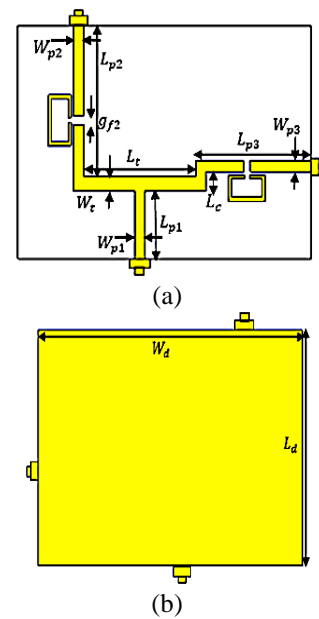


Figure. 13 The suggested diplexer's construction: (a) top view, and (b) Bottom view

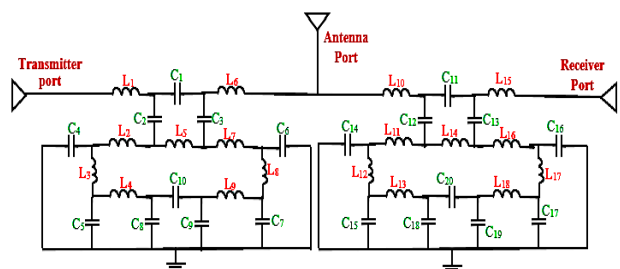


Figure. 14 The proposed BPF's LC equivalent circuit

loss of the filter. The band pass filter's unloaded quality factor  $Q_o$ , which can be calculated using Eq. (11), can serve as an indicator of its electrical performance and provide a better understanding of the band pass filter's insertion loss and selectivity, according to reference [30].

$$Q_L = \frac{1}{\frac{2}{Q_e} + \frac{1}{Q_o}} = \frac{f_0}{(\Delta f)_{3dB}} \quad (10)$$

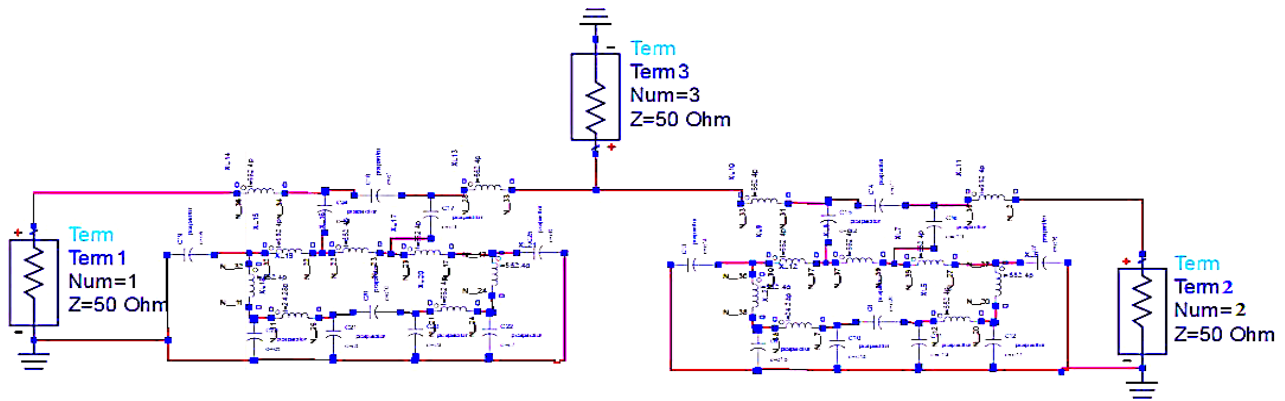


Figure. 15 The schematic circuit model of the proposed diplexer in ADS simulation program

Table 3. The suggested diplexer's dimensions

Parameter	Value (mm)	Parameter	Value (mm)
$W_d$	90	$W_{p2}$	3.065
$L_d$	70	$L_{p2}$	41.2
$W_{p1}$	3.056	$W_{p3}$	3.065
$L_{p1}$	25	$L_{p3}$	32
$W_t$	405	$g_{f2}$	2.5
$L_t$	33.9	$L_c$	5.4

$$Q_0 = \frac{Q_L}{(1-10^{-1L/20})} \quad (11)$$

With  $Q_L$  being the loaded  $Q$  determined by the equation, where  $f_0$  is the resonant frequency and  $(\Delta f)_{3dB}$  is the pass band's - 3 dB bandwidth (10).  $Q_e$  is an indicator of external quality.

The illustration in Fig. 10, demonstrates that the quality factor without any load increases as the separation between the two feed lines is widened. Expanding the space width from 2.3 mm to 4.1 mm results in larger filter dimensions, which in turn contribute to increased radiation loss and reduced conduction loss. This shift in current density, which is associated with the elevated quality factor without load, can be attributed to these observations, as noted in reference [30]. Fig. 11, depicts the proposed BPF as a fabricated prototype on a Rogers TMM4 substrate with dimensions of  $(25 \times 25) \text{ mm}^2$ . Fig. 12, illustrates a side-by-side evaluation of the measured and simulated S-parameters  $|S_{11}|$ , and  $|S_{21}|$  for the suggested BPF, employing the VNA spectrum. The simulation results closely aligned with the measurements, demonstrating a strong level of agreement.

## 2.2 The suggested microstrip diplexer's assembly

The two OLR-based BPFs discussed before and discussed in Sec. 2.1 are connected by a T-junction in this section to introduce the whole structure of the suggested diplexer as presented in Fig. 13. The

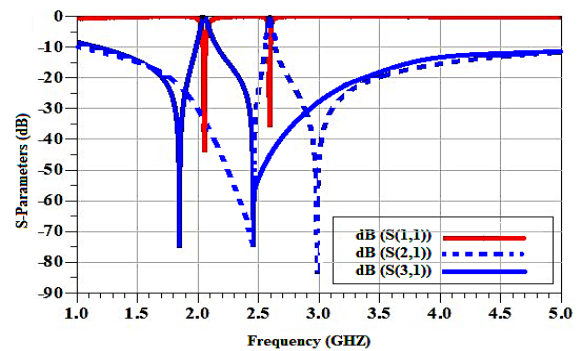


Figure. 16 Simulated scattering parameters of the LC equivalent circuit

isolation among transmit and receive channels is controlled by the width and length of the T-junction branches. Table 3 has a list of the proposed diplexer's dimensions.

A second method to studying this topology by converting the microstrip diplexer to its corresponding LC model, as shown in Fig. 14, in order to comprehend the behaviour of the suggested diplexer.

Within this LC representation, the characteristics  $L_{1.}, L_{2.}, L_{3.}, L_{4.}, L_{5.}, L_{6.}, L_{7.}, L_{8.}, L_{9.}, C_{1.}, C_{2.}, C_{3.}, C_{4.}, C_{5.}, C_{6.}, C_{7.}, C_{8.}, C_{9.}, C_{10}$  represent the transmitter BPF and  $L_{10.}, L_{11.}, L_{12.}, L_{13.}, L_{14.}, L_{15.}, L_{16.}, L_{17.}, L_{18.}, C_{11.}, C_{12.}, C_{13.}, C_{14.}, C_{15.}, C_{16.}, C_{17.}, C_{18.}, C_{19.},$  and  $C_{20}$  represent the Reciever BPF, the two filters grouped together with T junction. Fig. 15, represents the Schematic circuit model of the proposed diplexer in ADS simulation program, and Fig. 16, represents the simulated scattering parameters of the LC equivalent circuit. It is noticed that the behaviour of the diplexer is ideal and allows to pass frequencies 2.191GHz and 2.584 GHz at specified value of  $L$  and  $C$  as follow:  $C_1 = 0.18., C_2 = 0.16., C_3 = 7.8., C_4 = 0.12., C_5 = 0.3., C_6 = 0.11., C_7 = 9.0, C_8 = 0.1., C_9 = 0.14., C_{10} = 0.136., C_{11} = 0.2., C_{12} = 9.8., C_{13} = 3.2., C_{14} = 0.15., C_{15} = 7.2., C_{16} = 8.6., C_{17} = 9.2., C_{18} = 7.4., C_{19} = 0.174., C_{20} = 0.118.$  (all in PF),  $L_1 = 0.01., L_2 = 0.01., L_3 = 0.01.,$



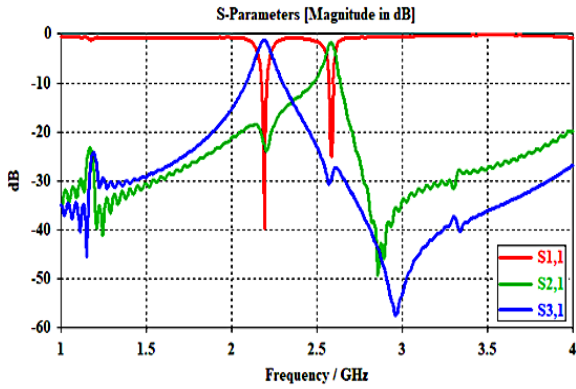


Figure. 17 The suggested diplexer's simulated S-parameters

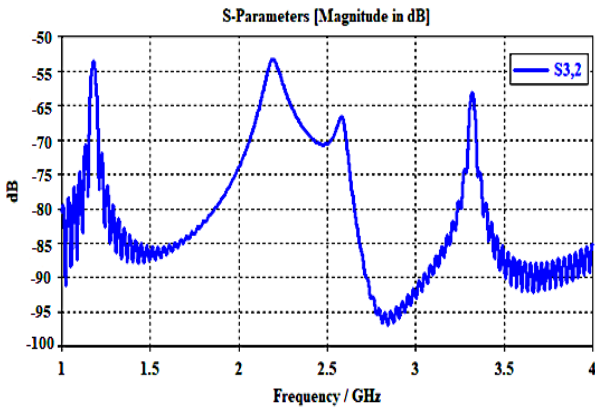


Figure. 18 The suggested diplexer's simulated scattering parameter  $|S_{32}|$

$L_4 = 0.01.$  ,  $L_5 = 0.01.$  ,  $L_6 = 0.01.$  ,  $L_7 = 0.009.$  ,  $L_8 = 0.01.$  ,  $L_9 = 0.01.$  ,  $L_{10} = 0.01.$  ,  $L_{11} = 0.01.$  ,  $L_{12} = 0.01.$  ,  $L_{13} = 0.01.$  ,  $L_{14} = 0.01.$  ,  $L_{15} = 0.01.$  ,  $L_{16} = 0.009.$  ,  $L_{17} = 0.01.$  ,  $L_{18} = 0.01.$  (all in nH).

Fig. 17, displays the simulation results of the proposed diplexer's scattering parameters  $|S_{11}|$ ,  $|S_{21}|$ , and  $|S_{31}|$  versus frequency. For the transmit and receive channels, respectively, the structure displays good insertion losses of around 1.2 dB and 1.7 dB, with fractional bandwidths of 1.24% at  $f_t = 2.191$  GHz and 0.636% at  $f_r = 2.584$  GHz. To put it another way, the transmit and receive bands' respective 3dB bandwidths are 36 MHz and 21.1 MHz, and the corresponding simulated return losses are 40 dB and 26 dB. The scattering parameter  $|S_{32}|$  is displayed in Fig. 18. The simulated isolation levels stand at 53.3 dB for 2.191 GHz and 66.5 dB for 2.584 GHz, respectively.

Additionally, the suggested diplexer achieves a modest frequency spacing ratio, denoted as R, equal to 0.1646. This measure is delineated as the difference between the transmit and receive frequencies  $\Delta f = |f_r - f_t|$  in relation to the central frequency  $f_c = (f_t + f_r)/2$  with R expressed as per

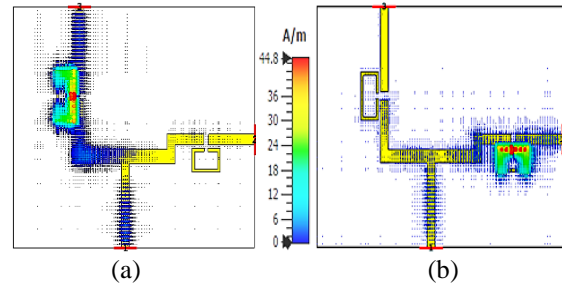


Figure. 19 Distribution of surface current in the suggested diplexer at: (a)  $f_t = 2.191$  GHz, and (b)  $f_r = 2.584$  GHz

the definition in reference [31].

$$R = \frac{\Delta f}{f_c} \tag{12}$$

Fig. 19, showcases the computed current distributions of the proposed diplexer at both the transmit and receive frequencies. When the diplexer is configured for transmission at  $f_t=2.191$  GHz, as seen in Fig. 19(a), there is a substantial current density along the path connecting port 1 to port 3, while the pathway from port 1 to port 2 remains inactive. In contrast, when the diplexer operates at the receive frequency  $f_r=2.584$  GHz, as depicted in Figure 19(b), the route from port 1 to port 2 exhibits significant current density, while the connection from port 1 to port 3 remains inactive. As a result, effective isolation is achieved between the broadcasting and receiving channels.

### 3. Constructed prototype of the suggested diplexer

The modelled diplexer, illustrated in Fig. 20, has been fabricated and printed on a Rogers TMM4 substrate with dimensions of  $(90 \times 70)$  mm<sup>2</sup>. The diplexer's performance was evaluated by measuring its S-parameters, namely  $|S_{11}|$ ,  $|S_{21}|$ , and  $|S_{31}|$ , using a Vector Network Analyzer (VNA) model Rohde & Schwarz ZVL20, as well as through simulations, as depicted in Fig. 21. The scattering parameter  $|S_{32}|$  of the diplexer was also compared between measurements and simulations in Fig. 22. Although there were slight differences in the frequency, the error percentage  $\frac{f_{measured} - f_{simulated}}{f_{measured}}$  did not exceed 5%.

### 4. Conclusion

The focus of this paper is on the development of a microstrip diplexer that utilizes OLR BPFs for wireless communication systems. The diplexer is engineered to attain excellent selectivity, robust isolation, minimal frequency spacing, and low

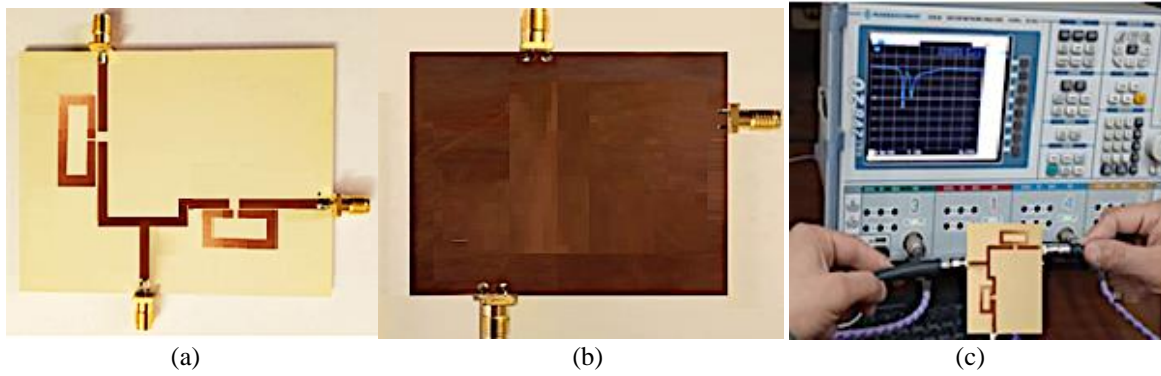


Figure. 20 Made-up model of the planned diplexer: (a) Overhead perspective, (b) Underneath perspective, and (c) Experimental arrangement

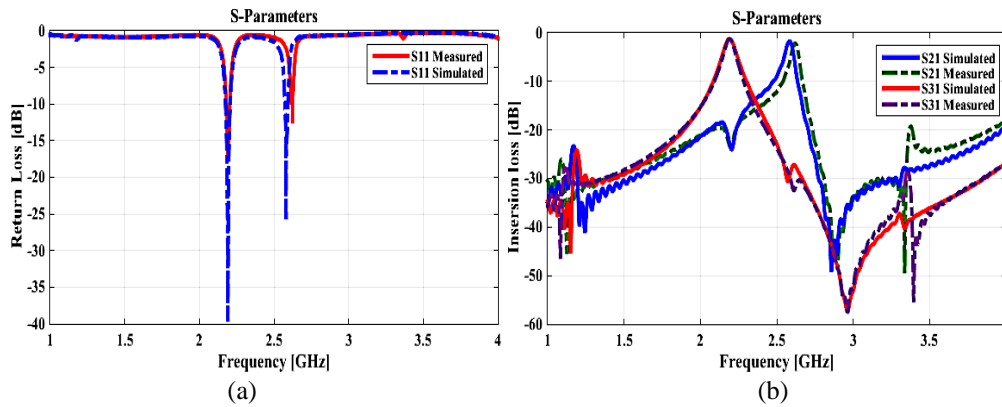


Figure. 21 The suggested diplexer's measured and simulated scattering parameters: (a)  $|S_{11}|$  and (b)  $|S_{21}|$ , and  $|S_{31}|$

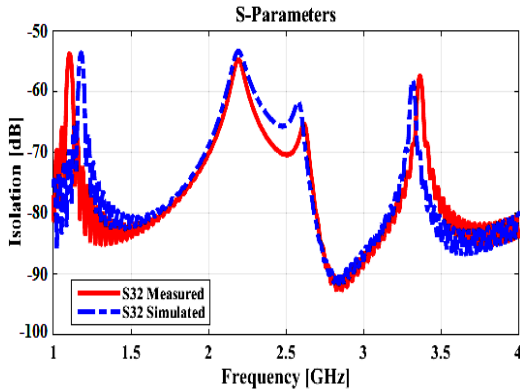


Figure. 22 The suggested diplexer's measured and simulated scattering parameter  $|S_{32}|$

insertion loss. It functions within the frequency range of  $f_t=2.191$  GHz and  $f_r=2.584$  GHz, employing the Roger TMM4 substrate known for its low electrical loss and stable dielectric constant across the frequency spectrum. The suggested diplexer design has been subjected to simulation using CST software, and the results demonstrate excellent performance with high isolation of approximately 53.3 dB and 66.5 dB for 2.191 GHz and 2.584 GHz, respectively. It also demonstrates the lowest insertion loss of approximately 1.2 dB for the transmit channel and 1.7 dB for the receive channel, as well as an acceptable frequency space ratio of approximately 0.1646 and

small fractional bandwidths of 1.24% and 0.636% for the transmit and receive channels, respectively. The performance of the manufactured diplexer was assessed, and a numerical equation for the LC equivalent circuit of the recommended BPF and the suggested microstrip diplexer was built. The finalized diplexer is well suited for wireless communication systems due to its miniaturized size, minimal insertion loss, and great selectivity.

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### Conflicts of interest

The authors declare no conflict of interest.

### Author contributions

Conceptualization, Rania Hamdy Elabd, Ahmed Jamal Abdullah Al-Gburi and Zahriladha Zakaria; methodology, Rania Hamdy Elabd; software, Ahmed Jamal Abdullah Al-Gburi; validation, Jamil Abedalrahim Jamil Alsayaydeh, Tale Saeidi, and Naba Jasim Mohammad; formal analysis, Jamil

Abdelrahim Jamil Alsayaydeh; investigation, Ahmed Jamal Abdullah Al-Gburi; resources, Rania Hamdy Elabd; data curation, Rania Hamdy Elabd; writing—original draft preparation, Rania Hamdy Elabd; writing—review and editing, Ahmed Jamal Abdullah Al-Gburi and Khaled Alhassoon; visualization, Naba Jasim Mohammad and Khaled Alhassoon; supervision, Ahmed Jamal Abdullah Al-Gburi; project administration, Ahmed Jamal Abdullah Al-Gburi; funding acquisition, Ahmed Jamal Abdullah Al-Gburi. All authors have read and approved the final manuscript.

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